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Modeling of Gapped Power Bus Structures for Isolation Using Cavity Modes

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Abstract
Power bus resonance characteristics of a gapped power-plane with a slit and a split power-plane with a gap were studied, using a fast algorithm based on a full cavity-mode resonator model and the segmentation method. Inductance and capacitance models were used to account for a field coupling along the slit and across the gap, respectively. Good agreements between the calculated and measured results were found to demonstrate the effectiveness and accuracy of our fast algorithm and the segmentation method, as well as the inductance model for the slit and the capacitance model for the gap.

Keywords
Power bus resonance, isolation, full cavity-mode resonator model, segmentation method, model for a slit, model for a gap.

1. INTRODUCTION

Based on the full cavity-mode resonator model [1], a fast algorithm can be applied for accurately calculating power bus resonance characteristics. The fast algorithm is attributed to a closed-form expression for the impedance Z-matrix of the power/ground planes, which is in the form of a singly infinite series [2-4]. This is assuming that the pattern of the power/ground planes is rectangular. When the pattern is entirely arbitrary, we must rely upon full wave numerical methods such as FDTD and FEM. However, in actual PCBs, an entirely arbitrary pattern of the power/ground planes is not very common; in many cases, the pattern consists of several segments, which have simpler shapes, such as rectangles. In addition, a slight difference in the circumference of the pattern does not affect the resonance characteristics very much, so that we may approximate its shape with some rectangles. In a previous work [5], the fast algorithm was extented for a pattern consisting of several segments of rectangles,

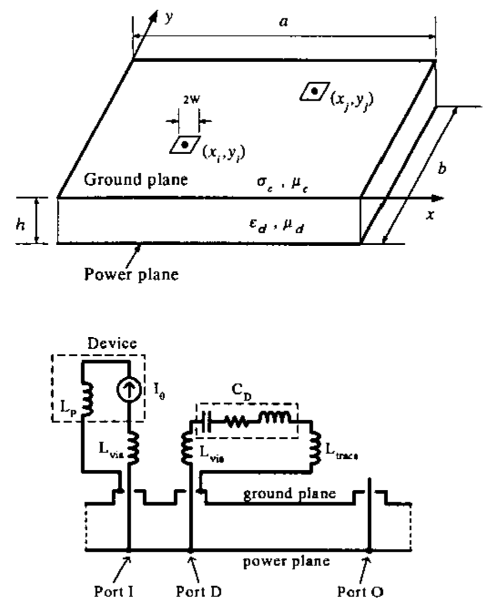

Fig.1: Geometry of rectangular power bus structure and its planar circuit model.

using the segmentation method [6,7] that was proposed many years ago for analyzing two-dimensional microwave planar circuits. The accuracy of the segmantation method has been demonstrated by good agreements between the calculated and measured results.

Board designers and EMC engineers frequently debate the merits of introducing finite sized gaps (slits) into the power and/or ground planes of multilayer PCBs. In fact, gapped power and/or ground planes are often used in high-speed digital circuit designs for minimizing the propagation of high-frequency noise on dc power buses. The high-frequency performance of gapped power-plane structures up to gigahertz range has been studed recently by using a full-wave approach called the CEMPIE [8] and by using a hybrid FEM/MOM approach [9]. In this presentation, we show that such gapped power-plane structures can also be analyzed by using the fast algorithm derived from the full cavity-mode resonator model, combined with the segmentation method and the coupling models for a slit and a gap.

2. FULL CAVITY-MODE RESONATOR MODEL

The full cavity-mode resonator model is an analytical description of the impedance matrix (Z-parameters) of an unloaded power/ground plane structure (a bare board). For a rectangular power/ground plane structure with length a and width b (see Fig. 1), an expression for quickly calculating the transfer impedance between the two ports on the power/ground planes was developed [3]:

$$Z_{ij} = \sum_{n=0}^{\infty} \frac{\omega \mu_d a h}{\mathrm{j}\, 2b} C_n \cos(k_{yn} y_i) \cos(k_{yn} y_j) \times \mathrm{sinc}^2(k_{yn} \mathrm{w}) \frac{[\cos(\alpha_n x_-) + \cos(\alpha_n x_+)]}{\alpha_n \sin \alpha_n}, \quad (1)$$

where $\mathrm{sinc}(x) = \sin(x)/x$, $k_{yn} = n\pi/b$, $\alpha_n = a\sqrt{\kappa^2 - k_{yn}^2}$, $x_\pm = 1 - (x_i \pm x_j)/a$, (x_i, y_i) and (x_j, y_j) are the coordinates of the center of the ith and jth ports in the x- and y-directions, respectively, w is much less than the wavelength of interest and represents the port (via) half width, h is the dielectric thickness between the power/ground planes, ω is the radian frequency, μ_d is the permeability of the dielectric, and $\mathrm{j} = \sqrt{-1}$. The constant C_n is assigned as $C_n = 1$ if $n = 0$, and $C_n = 2$ if $n \neq 0$. The complex transverse wavenumber κ is obtained as $\kappa^2 = \omega^2 \mu_d \varepsilon_d - \mathrm{j}2\omega\varepsilon_d Z_S/h$, where Z_S represents the surface impedance of the power/ground conductors. It should be pointed out that the dielectric loss naturally appears in the imaginary part of the dielectric constant, while the conductor loss is incorporated by the surface impedance of the conductors. When the board is loaded with passive components or active devices (Fig. 1), the impedance characteristics of the loaded board can be analyzed by considering the loaded board as a multi-port circuit network interconnected by the Z-matrix elements of the corresponding bare board [3, 4].

The aforementioned expression for the rectanglar power/ground plane can be easily applied to those geometries that result from the connection of rectangles by using the segmentation method [5-7]. In the method, the interconnection between segments

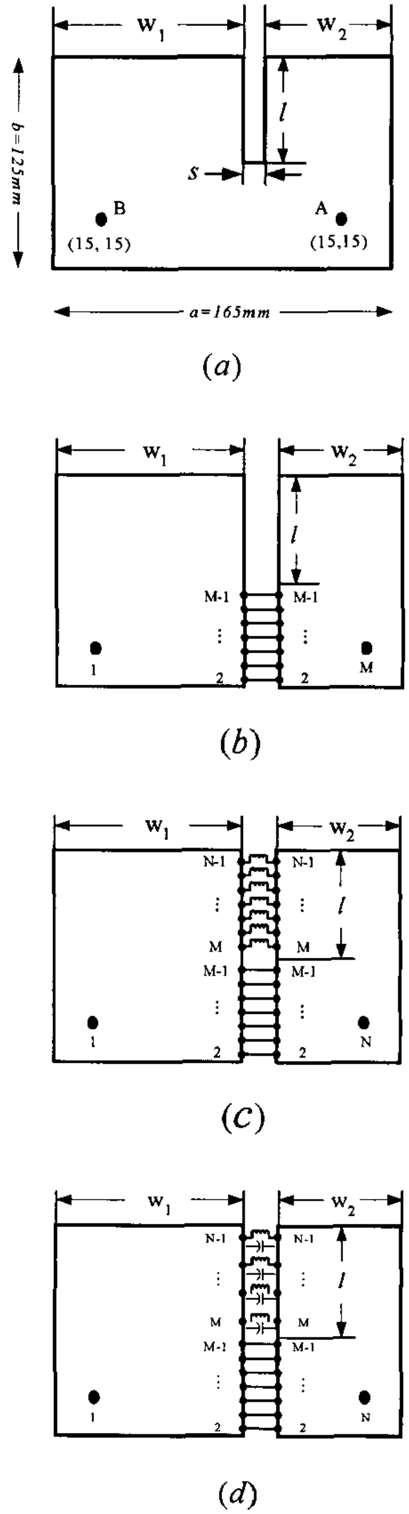

Fig.2: Geometry of gapped power-plane with a slit and its analytical models.

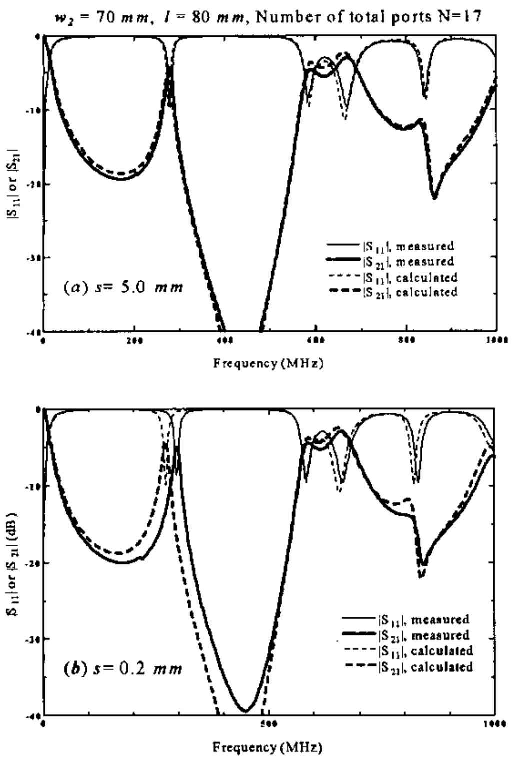

Fig.3: Comparisons between calculated and measured $|S_{11}|$ and $|S_{21}|$ for the board of Fig. 2(a) with analytical model of Fig. 2(b).

is discretized into a finite number of virtual ports, in which continuous voltage and current distributions along the interconnection are approximated by stepped functions. The number of ports used for the interconnection therefore determines the accuracy of the segmentation method.

3. MODELS FOR A SLIT AND A GAP

To prevent noise propagating from its source to susceptible devices, the gapped power plane has been used as a technique to mitigate noise in digital circuit design. EMC engineers have also developed many "split" or "gapped" ground plane design strategies for reducing the common mode radiation from cables attached to the PCBs. The RF performance of such segmented structures up to the gigahertz frequency range requires, in principle, a full-wave approach like the CEMPIE [8] or the hybrid FEM/MOM approach [9]. In this section, we show that the resonance characteristics of a gapped power bus can also be analyzed by using the full cavity-mode resonator model combined with the segmentation method. However, directly applying the segmentation method for the problem fails to give correct results, due to the fact that the field coupling between the adjacent edges of the segments along a slit or across a gap has not been taken into account in the full cavity-mode resonator model. Furthermore, as we show later, the model used to account for the coupling effect for a slit is different from that for a gap; thus these two cases are discussed separately.

3.1 A Power-Plane with a Slit

We at first consider a power-plane with a slit shown in Fig. 2(a). When the slit is wide, we may treat the whole pattern as a combination of three rectangular segments. When the slit is narrow, assuming that the slit width s is small enough so that we may ignore the electric length of s, directly applying the segmentation method then leads to the analytical model shown in Fig. 2(b) where the two segments are interconnected by the virtual ports at the part with a conducting bridge and are opened at the part with a slit. The $(M-2)$ virtual ports are uniformly distributed over the conducting bridge with a port width $\mathrm{w} = (b-l)/(M-2)$ and are assigned numbers ranging from 2 to $m = (M-1)$. The observing ports B on the left segment and A on the right are assigned numbers 1 and M. The formulation of the multi-port circuit analysis for the structure can be found in an earlier paper[10].

With the model in Fig. 2(b), the calculated $|S_{11}|$ at port A and $|S_{21}|$ from port A to port B were compared with the measured ones in Fig. 3 for a slit length $l = 80$ mm and the location $w_2 = 70$ mm with the widths (a) $s = 5.0$ mm and (b) $s = 0.2$ mm, respectively. The dielectric permittivity was assumed to be a constant $\varepsilon_r = 4.3 - \mathrm{j}\,0.086$ (material: FR-4 glass epoxy) for all frequencies. As shown in Fig. 3, very good agreements were obtained for a wide slit (Fig. 3(a), $s = 5.0\,\mathrm{mm} \simeq 3\,h$, h: the dielectric thickness) between the calculated and measured results while the obvious differences were observed for a narrow slit (Fig. 3(b), $s = 0.2$ mm) at lower-order resonances. The reason for this is attributed to the field coupling between the adjacent edges of the segments along the slit when the width is narrow, because such coupling had not been taken into account in the full cavity-mode resonator model due to the "open" sidewalls (the perfect magnetic walls) used as the boundaries for the segments.

When a slit is cut on the power-plane, it cuts off the path of the electric current for the modes whose currents flow toward a direction perpendicular to the slit. The current has to flow around the slit, and its path gets lengthens. When the slit length l is less than one fourth of the wavelength ($l < \lambda/4$), the effect can be described in terms of an equivalent

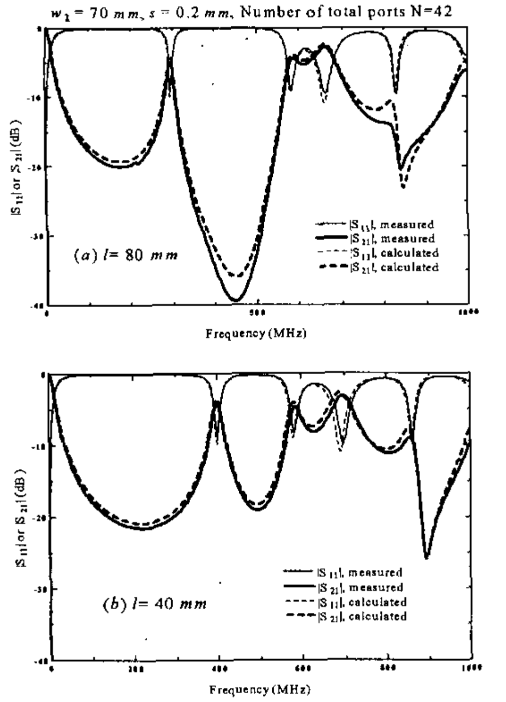

Fig.4: Comparisons between calculated and measured $|S_{11}|$ and $|S_{21}|$ for the board of Fig. 2(a) with analytical model of Fig. 2(c).

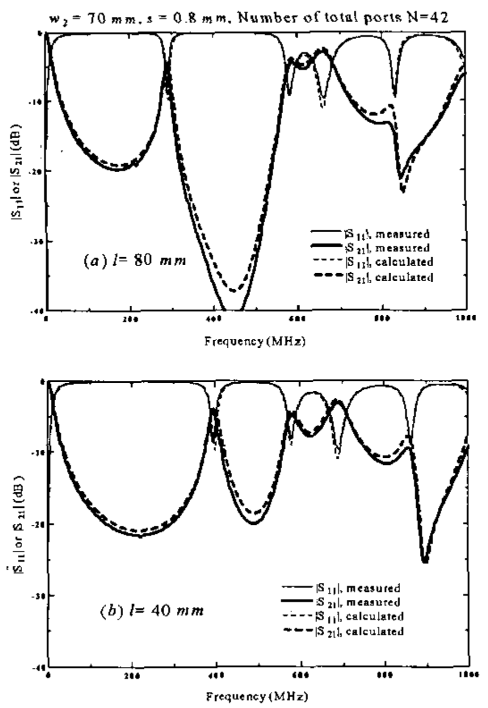

Fig.5: Same as Fig. 4 but for slit width $s = 0.8$ mm.

series inductance that is distributed along the slit as a multiport network, as shown in Fig. 2(c). The total partial inductance of the slit can be determined by visualizing it as a shorted coplanar slot-line with half the length of the slit. When the coplanar slot-line is symmetrical, the expression for this inductance can then be summarized as [11]

$$L = \frac{\pi \mu_0 l}{2 \ln \left[2 \frac{1+\sqrt{k'}}{1-\sqrt{k'}} \right]} \quad \text{(Henries)} \tag{2}$$

with

$$k' = \sqrt{1-k^2}, \quad k = s/(s+2w), \tag{3}$$

where w is the width of the two strip lines and is approximated by the smaller of w_1 and w_2 in Fig. 2(a) for our present asymmetrical case. This inductance is then distributed along the slit as part of a multiport network, and the port inductance increases as the distance from this port to the shorted conducting bridge increases. Arranging the port numbers for the distributed inductance as M to $(N-1)$, as shown in Fig. 2(c), the port inductance is given as

$$L_p = \frac{L}{2} \frac{(N-M)(N-M+1)}{(N-p)}, \tag{4}$$

where $p = M, M+1, \cdots, N-2, N-1$. The calculated $|S_{11}|$ at port A and $|S_{21}|$ to port B were again compared with the measured ones in Figs. 4 and 5, for different values of the slit width and length. The effectiveness of the inductance model for the slit was demonstrated by good agreements between the calculated and measured results for all of these four cases.

As already described, we accounted for the presence of a slit on the power plane by using an inductance. However, one could argue that capacitive coupling effect across the narrow slit could be present as well. Actually, when the power plane is completely split into two segments by a narrow gap, the coupling between the two segments by a gap capacitance on the order of several picofarads is very important for studying the isolation effect between these two segments provided by the gap, as we will show in the next subsection. With Eq. (5) given below for this gap capacitance, we calculated the $|S_{11}|$ and $|S_{21}|$ for our case of having a conducting bridge between the two segments. Both the slit inductance and the gap capacitance were considered. The results plotted in Fig. 6 were compared with the ones obtained when the slit is modeled as open or is modeled by only the slit inductance. The overlap found between two curves of only L or both L and C at low frequencies below 1 GHz indicates that the gap capacitance

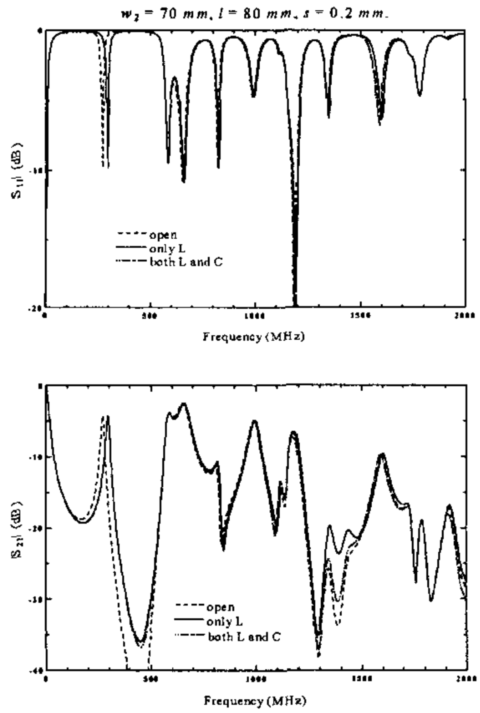

Fig.6: Comparisons of calculated $|S_{11}|$ and $|S_{21}|$ with different models for slit.

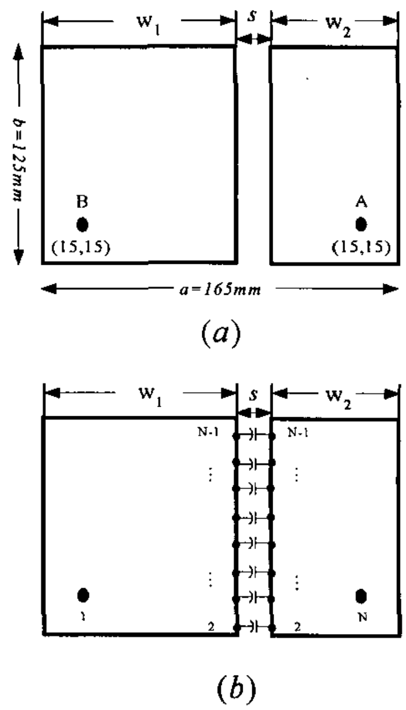

Fig.7: Geometry of split power-plane with a gap and its analytical model.

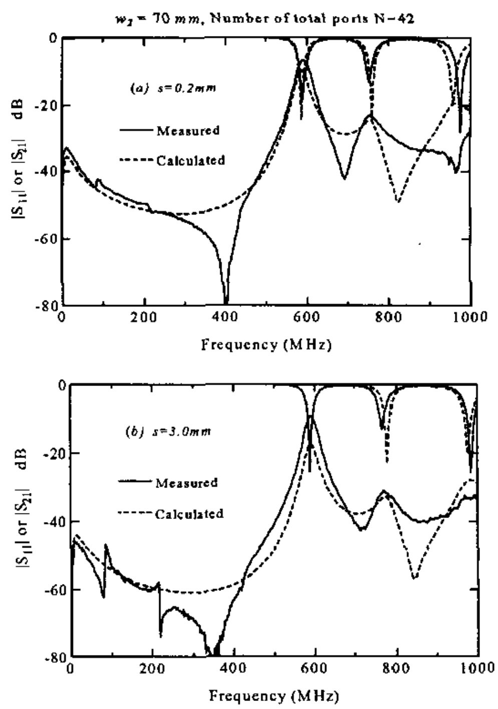

Fig.8: Comparisons between calculated and measured $|S_{11}|$ and $|S_{21}|$ for the board of Fig. 7(a) with analytical model of Fig. 7(b).

has little effect when the slit is shorted by a conducting bridge, while the slight differences among the three curves at high frequencies above 1 GHz indicate that whichever model is used for the slit is no longer important as the frequency increases.

3.2 A Split Power-Plane with a Gap

We consider a split power-plane with a gap shown in Fig. 7(a) next. The effect of split power island structures on power bus isolation has been experimentally and theoretically investigated [8, 9, 12]. Without any direct connection between the two power segments or islands, the primary coupling mechanism is capacitive coupling across the gap. The gap capacitance may be obtained by considering the two segments as parallel-coupled microstrip lines [13]. It is given as

$$C = \frac{\varepsilon_0 \varepsilon_r b}{\pi} \ln\left[\coth\left(\frac{\pi}{4}\frac{s}{h}\right)\right] + \frac{\varepsilon_0 b}{\pi} \ln\left[2\frac{1+\sqrt{k'}}{1-\sqrt{k'}}\right] \quad \text{(Farads)}, \qquad (5)$$

where s is the gap width, and k' is defined by Eq. (3). With this capacitance and the model of Fig. 7(b), the calculated $|S_{11}|$ at port A and $|S_{21}|$ to port

B were compared with the measured ones in Fig. 8, for gap widths $s = 0.2$ mm and $s = 3.0$ mm, respectively. The agreement between the calculated results and the measured ones is good, except for the frequencies where both power islands have degenerated resonance due to the same width b. Rather than using the capacitance model, greater accuracy in the numerical simulation can be achieved by using a multiport mutual coupling network [14] representing the coupling between the two adjacent segments. However, a great deal of computational effort has to be made due to the increase in the number of ports and due to the need to calculate the mutual coupling coefficients.

4. CONCLUSION

A fast algorithm based on the full cavity-mode resonator model was applied for analyzing power bus resonance characteristics of gapped power-plane structures, combined with the segmentation method. For a power-plane with a slit, an inductance model was introduced for representing the coupling effect due to the slit. The value of the inductance is obtained by treating the slit as a shorted coplanar slot-line. For a split power-plane with a gap, a capacitance model is used for taking into account the field coupling across the gap. The value of the capacitance is determined by considering two segments of the split power-plane as parallel-coupled microstrip lines. Good agreements between the calculated and measured results demonstrated the effectiveness and accuracy of our fast algorithm and the segmentation method, as well as the inductance model for the slit and the capacitance model for the gap.

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